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[54] **SEMICONDUCTOR CIRCUITS FOR GENERATING REFERENCE POTENTIALS WITH PREDICTABLE TEMPERATURE COEFFICIENTS**

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[51] **Int. Cl.**² **G05F 1/58**

[52] **U.S. Cl.** .... **323/19**; 307/297; 323/22 T; 323/23

[58] **Field of Search** 323/1, 4, 9, 19, 227, 323/68; 307/296, 297; 330/32

[56] **References Cited**

U.S. PATENT DOCUMENTS

3,579,133	5/1971	Harford	330/32
3,617,859	11/1971	Dobkin et al.	323/4
3,659,121	4/1972	Frederiksen	307/297
3,781,648	12/1973	Owens	323/4
3,794,861	2/1974	Bernacchi	307/297
3,887,863	6/1975	Brokaw	323/22 T

OTHER PUBLICATIONS

"Stable Voltage Ref. Crt." by J. E. Gersbach, IBM Tech. Disc. Bull., vol. 18, No. 7, Dec. 1975, pp. 2091–2092.

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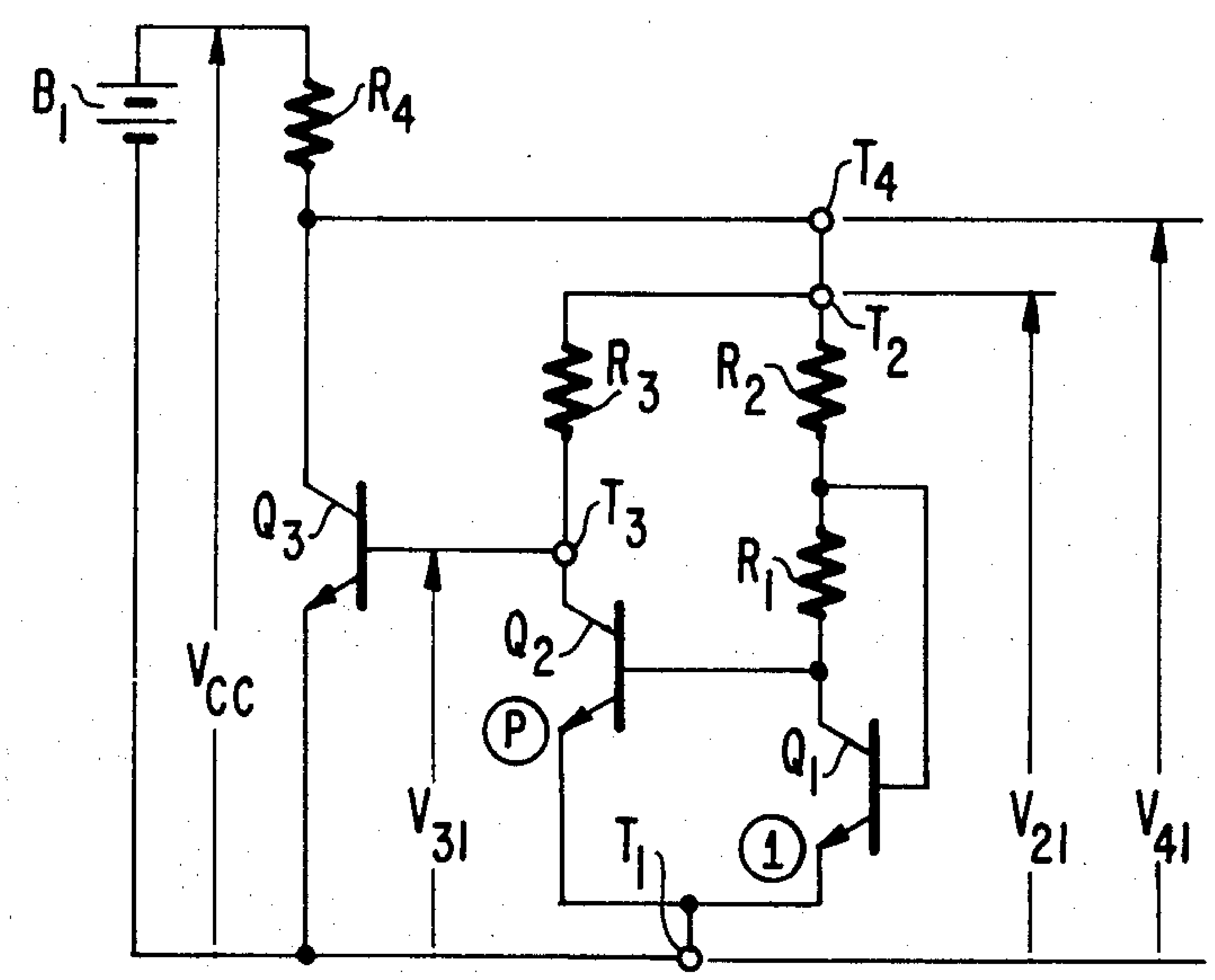
[57] **ABSTRACT**

A positive-temperature-coefficient difference between the emitter-to-base potentials of two transistors in particular configuration is scaled up and added to one of the emitter-to-base potentials to develop a potential, a multiple of which is supplied as the reference potential.

7 Claims, 7 Drawing Figures



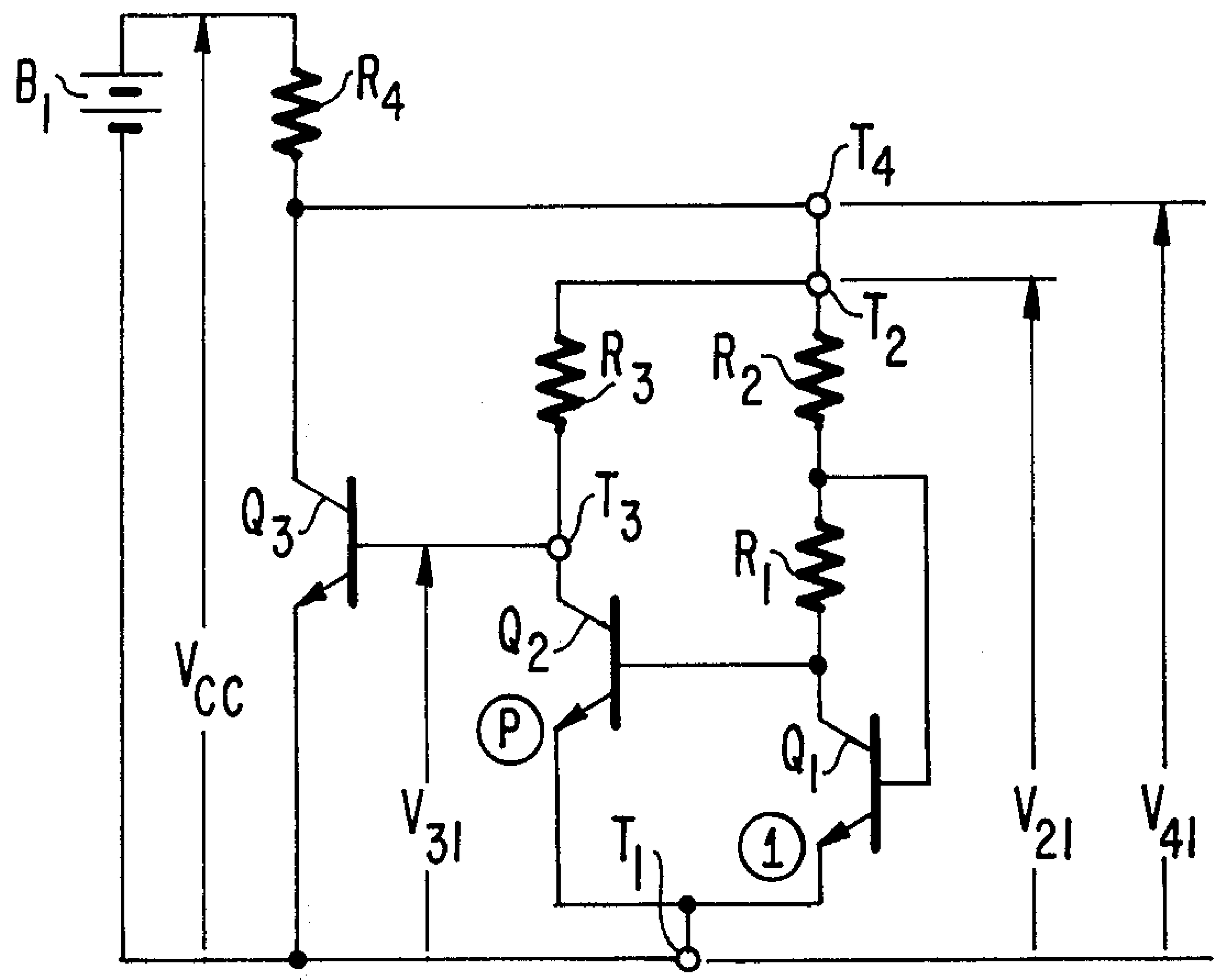


Fig. 1.

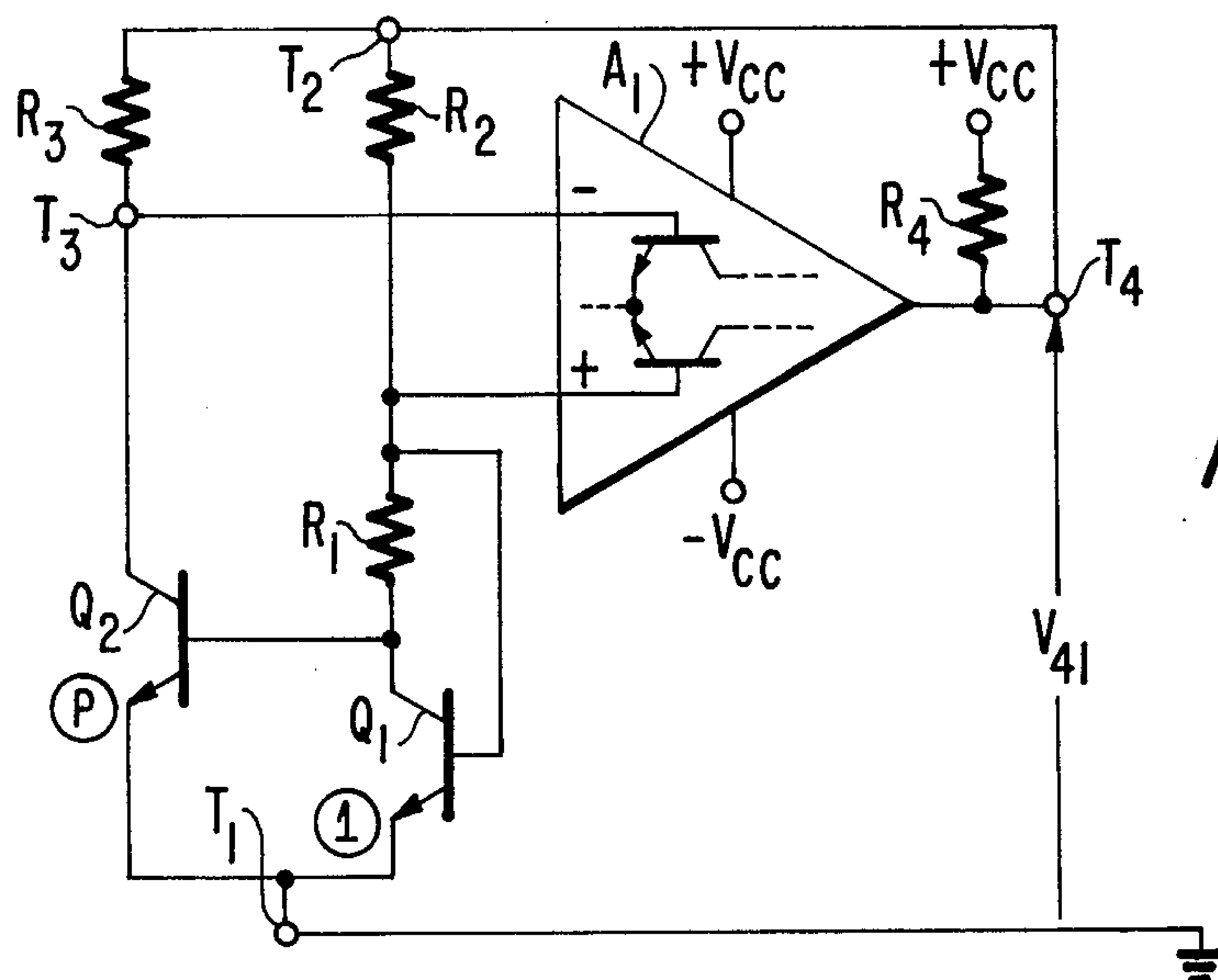


Fig. 2.

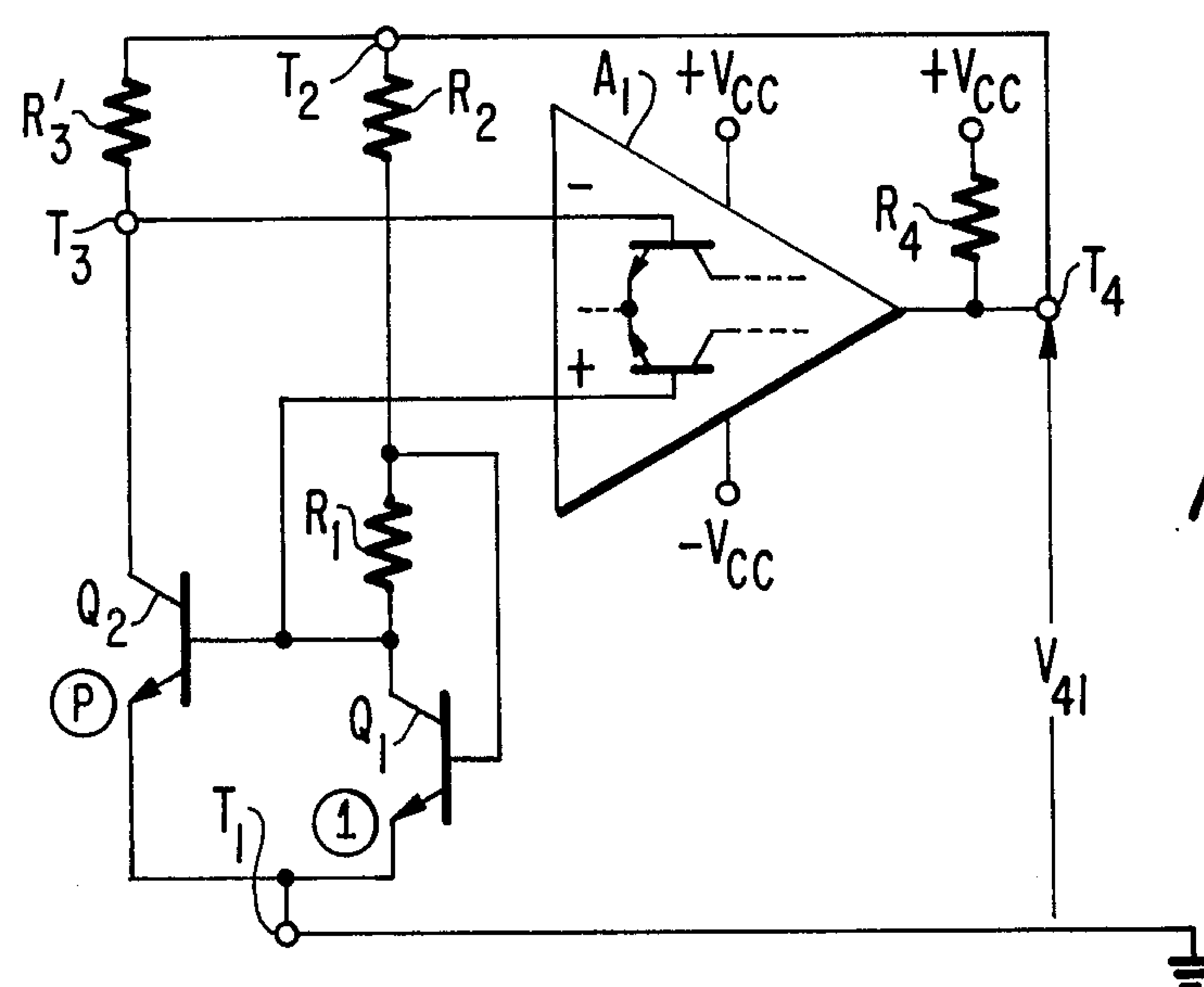


Fig. 3.

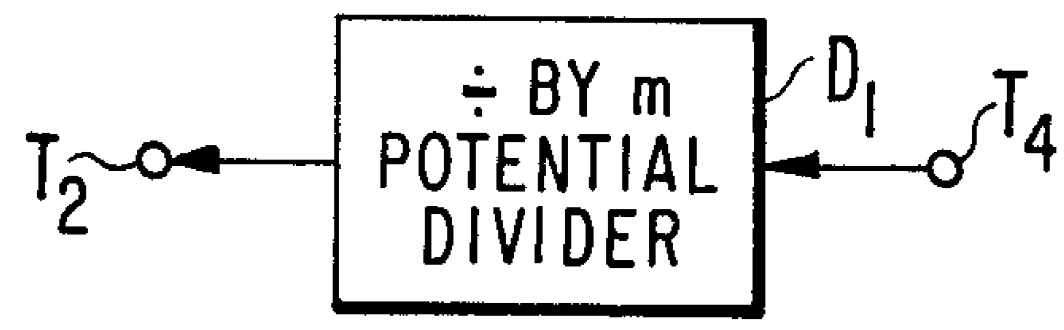


Fig. 4.

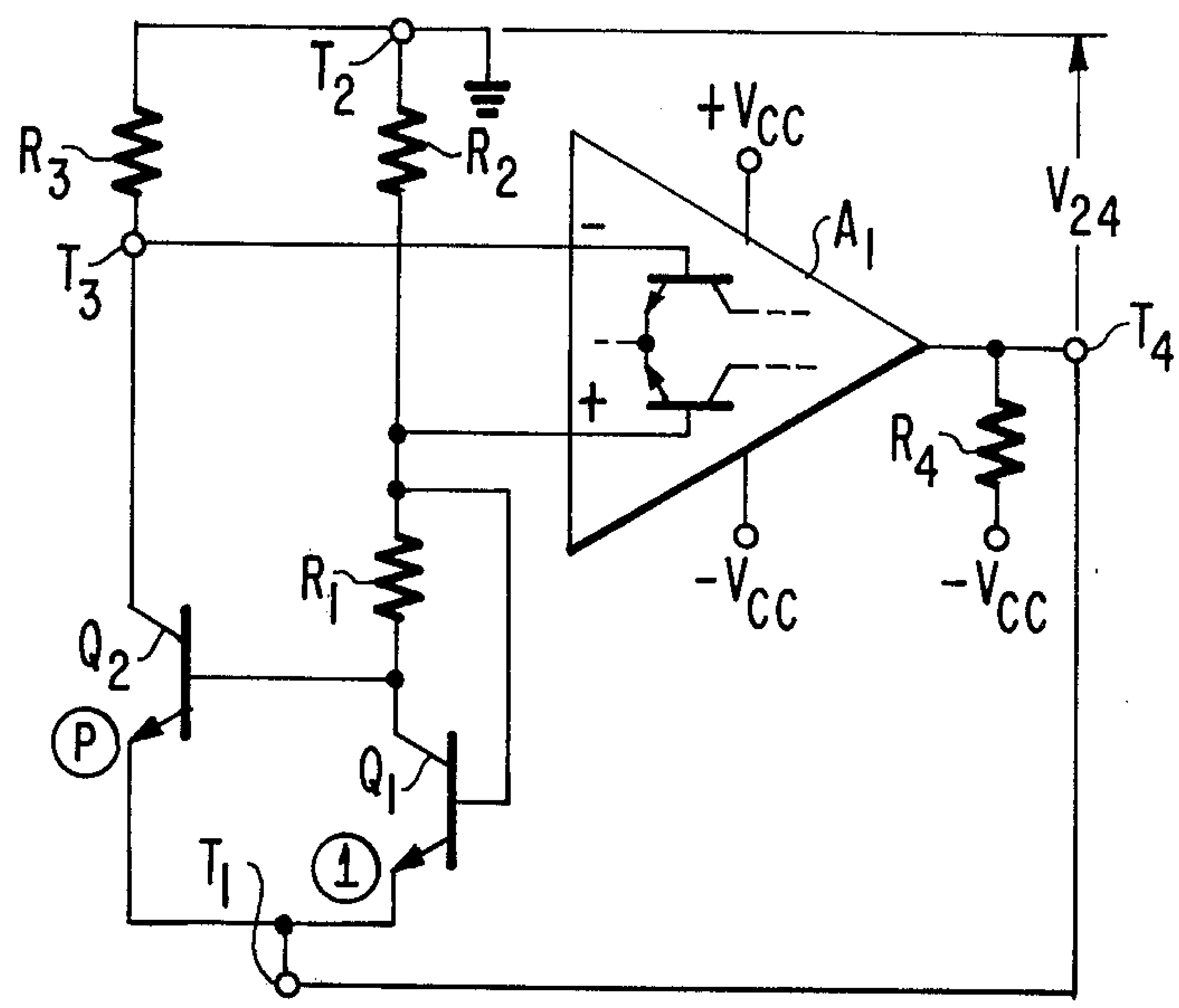


Fig. 5.

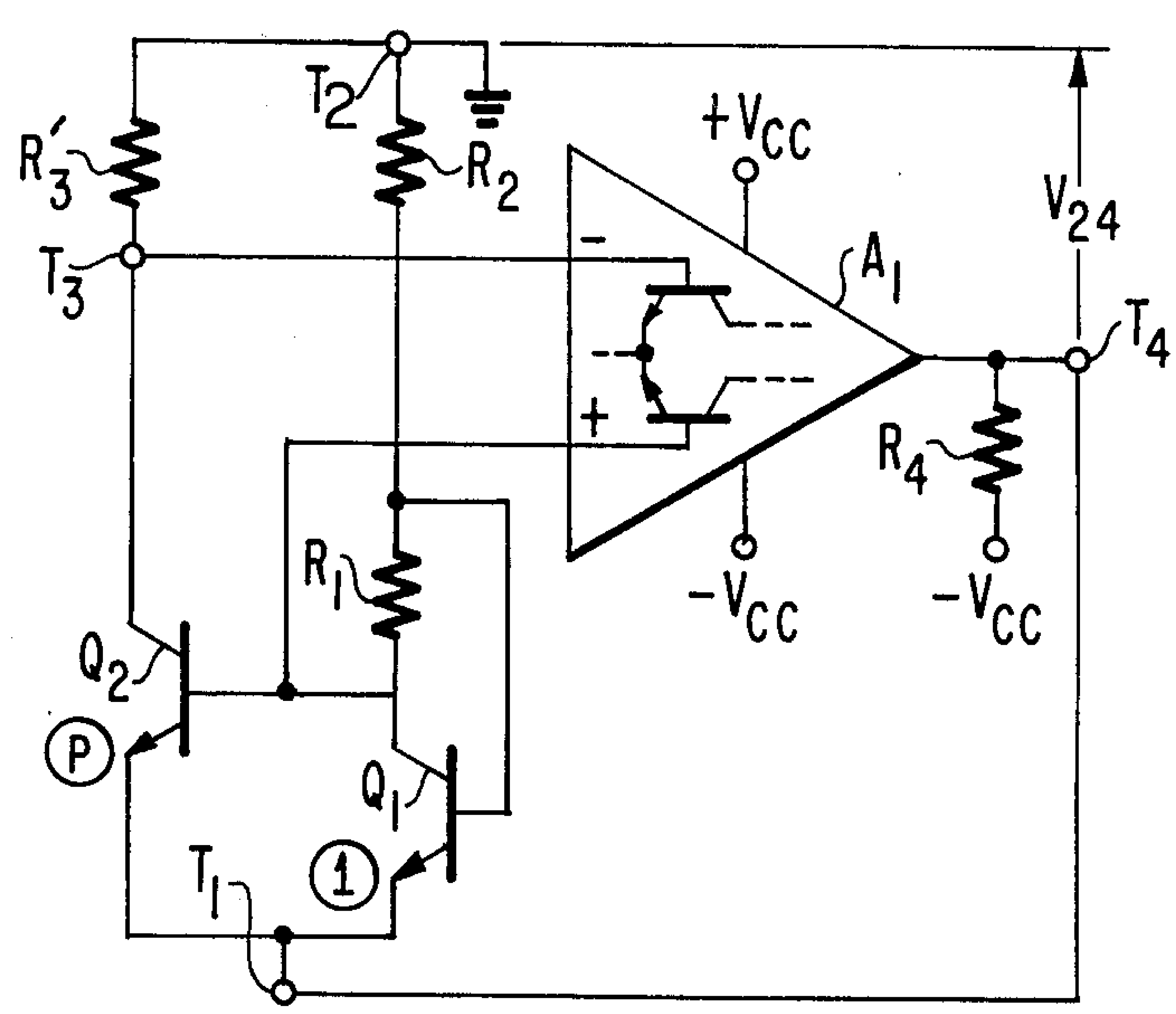


Fig. 6.

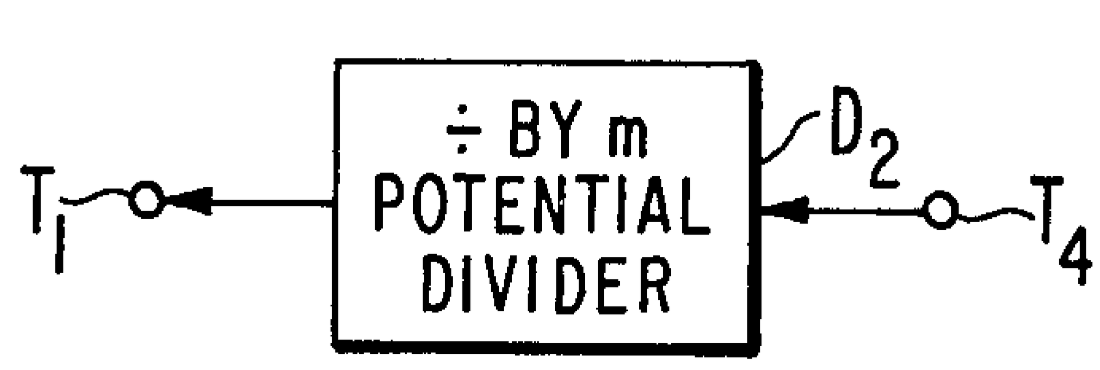


Fig. 7.

SEMICONDUCTOR CIRCUITS FOR GENERATING REFERENCE POTENTIALS WITH PREDICTABLE TEMPERATURE COEFFICIENTS

Circuits are known for generating reference potentials related to $V_{g(0)}$, the band-gap potential of a semiconductor material such as silicon, extrapolated to zero Kelvin. They may be particularly suited to fabrication in integrated circuit form. See R. J. Widlar's article, "New Developments in IC Voltage Regulators" appearing on pp. 2–7 of *IEEE Journal of Solid State Circuits,* Vol. SC-6, No. 1, February 1971, and K. E. Kuijk's article "A Precision Reference Voltage Source" appearing on pp. 222–226 of IEEE Journal of Solid State Circuits, Vol. SC-8, No. 3, June 1973. See, too, U.S. Pat. Nos. 3,271,660 (Hilbiber), 3,617,859 (Dobkin et al.), 3,648,153 (Graf) and 3,887,863 (Brokaw).

The present invention is embodied in a reference potential generator with superior potential regulation properties. While not restricted thereto, a number of embodiments of the invention are suitable for generating potentials related to $V_{g(0)}$.

In the drawing:

each of FIGS. **1, 2, 3, 5** and **6** is a schematic diagram of a reference potential generator furnishing a reference potential substantially equal to the $V_{g(0)}$ of the semiconductive material from which its transistors are fabricated;

FIG. **4** is a block schematic diagram showing how the circuits of FIGS. **1, 2** and **3** may be modified to increase the reference potential by a factor *m;* and

FIG. **7** is a block schematic diagram showing how the circuits of FIGS. **5** and **6** may be modified to increase the reference potential by a factor *m.*

Each of the FIGS. **1, 2, 3, 5** and **6** includes first and second transistors Q_1 and Q_2, respectively, and first, second and third resistive elements R_1, R_2 and R_3, respectively. Each also includes first, second and third terminals T_1, T_2 and T_3, respectively. Q_1 and Q_2 are operated at the same absolute temperature T expressed in units Kelvin. Q_1 and Q_2 have respective base-emitter junctions with similar profiles and respective effective areas in *l:p* ratio, *p* being a positive number, as indicated by the encircled numbers near their respective emitter electrodes.

In FIG. **1**, a bias means comprising the series connection of battery B_1 supplying potential V_{CC} and resistor R_4 tends to keep terminal T_4 (and terminal T_2 connected thereto) at a different potential from terminal T_1. A degenerative feedback connection is provided wherein V_{21}, the difference in potential between T_1 and T_2, is coupled via R_3 to terminal T_3 at the base electrode of transistor Q_3. The feedback biases Q_3, which has its emitter electrode connected to T_1, into conduction. The resultant collector-to-emitter current demand presented by Q_3 is met from battery B_1, with the collector current I_{CQ3} of Q_3 causing a potential drop across R_4 that reduces the potential V_{41} between T_1 and T_4 to carry out shunt potential regulation of V_{21}. This degenerative feedback connection would—were the connection comprising Q_1, Q_2, R_1 and R_2 not present—operate to reduce V_{21} to a value equal to the emitter-to-base potential V_{BEQ3} of Q_3 required to support a collector current flow substantially equal to $(V_{CC} - V_{BEQ3})/R_4$—e.g., somewhere from 500 to 700 millivolts.

The connection comprising elements Q_1, Q_2, R_1 and R_2 provides for a regenerative feedback connection in addition to the degenerative feedback connection described. At low values of V_{21}, the regenerative feedback connection has sufficient gain to overwhelm the effects of the degenerative feedback connection. But as V_{21} is increased, the gain of the regenerative feedback connection is reduced, and at some predictable value of V_{21}, the degenerative and regenerative feedback connections are so proportioned that the Nyquist criterion for stable equilibrium is met.

At low values of V_{21}, very little current will flow through the series combination of R_2 and Q_1 (regarded as a self-biased transistor). The portion of this current flowing through R_1 will cause a negligibly small potential drop across R_1, so the emitter-to-base potentials of Q_1 and Q_2 will be substantially equal. Current mirror amplifier action will thus obtain between transistors Q_1 and Q_2. The collector current I_{CQ2} of Q_2 will accordingly be about *p* times as large as the collector current I_{CQ1} of Q_1, the major component of the current flowing through the series combination of R_2 and Q_1 (regarded as a self-biased transistor). Any increase of V_{21} above V_{BEQ1} will cause a current $(V_{21} - V_{BEQ1})/R_2$ to flow through R_2, the major portion of which current will flow as I_{CQ1}. I_{CQ2} will be about *p* times as large as I_{CQ1} —i.e., $p(V_{21} - V_{BEQ1})/R_2$— causing a potential drop V_{32} across R_3 substantially equal to $p(V_{21} - V_{BEQ1})R_3/R_2$. So, if pR_3/R_2 be substantially larger than unity, increasing V_{21} will decrease rather than increase the potential V_{31} appearing between terminals T_1 and T_3 and applied as base-emitter potential to Q_3. Conduction of Q_3 will be suppressed, permitting V_{21} to grow towards its upper limit value of V_{CC}.

At higher values of V_{21}, the current $(V_{21} - V_{BEQ1})/R_2$ through R_2 increases. The major portion of this current flows as I_{CQ1} through R_1 to cause a potential drop across R_1. For each 18 millivolts of drop across R_1, I_{CQ2} is reduced by an additional factor of two compared to I_{CQ1}. So, while I_{CQ2} as well as I_{CQ1} increases with increasing V_{21}, its increase is slower than that of I_{CQ1}. I_{CQ1} increases almost linearly with increasing V_{21}, and it will be shown that I_{CQ2} increases substantially less than linearly with increasing V_{21}. The current flowing from T_2 to T_3 via R_3 has a value $(V_{21} - V_{BEQ3})/R_3$ and so increases substantially linearly with increasing V_{21}, at some value of V_{21} overtaking I_{CQ2} in amplitude sufficiently to provide substantial base current to Q_3. This base current renders Q_3 conductive to carry out shunt regulation of V_{21} against further increase.

Consider now why I_{CQ2} increases substantially less than linearly with increasing V_{21}. The operation of transistors Q_1 and Q_2 can be expressed in terms of the following expressions, as is well-known.

$$V_{BEQ1} = (kT/q)ln(I_{CQ1}/A_{Q1}J_S) \quad (1)$$

$$V_{BEQ2} = (kT/q)ln(I_{CQ2}/A_{Q2}J_S) \quad (2)$$

where V_{BEQ1} and V_{BEQ2} are the respective base-emitter junction potentials of Q_1 and of Q_2, k is Boltzmann's constant, T is the absolute temperature at which Q_1 and Q_2 are both operated, q is the charge on an electron, I_{CQ1} and I_{CQ2} are the respective collector currents of Q_1 and of Q_2, A_{Q1} and A_{Q2} are the respective effective areas of the base-emitter junctions of Q_1 and Q_2, and J_S is a saturation current density term presumed to be common to Q_1 and Q_2. At lower levels of input current applied to terminal T_4, the collector current of Q_1 is commensurately low, so that the base potential of Q_1 is applied to the base electrode of Q_2, without substantial drop across resistance R_1 due to I_{CQ1}. Eliminating V_{BE} between equations 1 and 2, I_{CQ2}/I_{CQ1} at very low levels of collector current can be shown to be as follows:

$$(I_{CQ2}/I_{CQ1}) = A_{Q2}/A_{Q1} = p \tag{3}$$

With increasing level of the input current, which I_{CQ1} is adjusted to equal, the drop V_1 across resistor R_1, essentially equal to $I_{CQ1}R_1$, is increased.

$$V_1 = V_{BEQ1} - V_{BEQ2} \tag{4}$$

Substituting equations **1**, **2** and **3**, into equation **4**, yields the following expression.

$$(I_{CQ2}/I_{CQ1}) = p\,\exp^{-1}(qV_1/kT) \tag{5}$$

The potential drop V_2 across R_2 is caused primarily by the flow of I_{CQ1} and is equal to the difference between V_{21} and V_{BEQ1}.

$$V_2 = I_{CQ1}R_2 \tag{6}$$

$$V_2 = V_{21} - V_{BEQ3} \tag{7}$$

An expression for I_{CQ1} can be obtained by cross-solving equations **6** and **7**.

$$I_{CQ1} = (V_{21} - V_{BEQ3})/R_2 \tag{8}$$

V_1 is caused primarily by the flow of I_{CQ1}.

$$V_1 = I_{CQ1}R_1 \tag{9}$$

Substituting equations 8 and 9 into equation 5, one obtains equation **10** describing I_{CQ2} in terms of V_{21}.

$$I_{CQ2} = p(V_{21} - V_{BEQ3})/R_2\exp(R_1/R_2)(V_{21} - V_{BEQ3})(q/kT) \tag{10}$$

The improved regulation characteristics of the reference potential generators built in accordance with the present invention are due to the very great percentage change in the current gain of the configuration comprising elements Q_1, Q_2, R_1 and R_2 and linking T_2 to T_3 to apply non-linear regenerative collector-to-base feedback to Q_3, responsive to small percentage changes in V_{21}. This percentage change in current gain with small percentage change in V_{21} is substantially superior to the non-linear regenerative feedback configuration as used by Widlar and Brokaw, differing from that shown by R_1 being replaced by direct connection and by the emitter of Q_2 being provided an emitter degeneration resistance. The current amplifier comprising elements Q_1, Q_2, R_1 and R_2 is per se known from U.S. Pat. Nos. 3,579,133 (Harford) and 3,659,121 (Frederiksen), but its non-linear current gain properties are not made use of as in the present invention.

Consider now how V_{21} may be regulated to be substantially equal to $V_{g(0)}$ the bandgap potential, as extrapolated to zero Kelvin, of the semiconductor material from which Q_1, Q_2 and Q_3 are made. $V_{g(0)}$ exhibits zero temperature coefficient and, assuming the transistors to be silicon transistors, has a value of about 1.2 volts. One can discern that the FIG. **1** reference potential generator is capable of synthesizing $V_{g(0)}$ since V_{21} is equal to the sum of the base-emitter offset potential of a transistor (Q_1) and a potential proportional to the difference in the base-emitter potentials of two transistors (the drop across R_2), such a summation being a known technique for synthesizing $V_{g(0)}$. The potential drop across R_2 is proportional to the drop across R_1 since: R_1 and R_2 conduct substantially the same current, and the drop across R_1 is known to equal $V_{BEQ1} - V_{BEQ2}$.

Knowing V_{CC} and what V_{41} is to be in terms of $V_{g(0)}$, one can select a value of R_4 in accordance with Ohm's Law to provide a convenient nominal value of operating current, respective portions of which flow to Q_3 as collector current I_{CQ3}, through R_3, and through the series combination of R_2 and self-biased Q_1. V_{21} will have a value substantially equal to 1236mV and V_{BEQ1} is about 550 - 700mV depending on I_{CQ1}. So the potential drop V_2 across R_2 is about 540 - 690mV. R_2 can be calculated by Ohm's Law, dividing the 540 - 690mV drop by I_{CQ1}. The potential drop V_1 across R_1 is typically chosen to be 60mV or so at equilibrium, so the scaling factor between R_1 and R_2 is not too large, this drop divided by I_{CQ1} yields a value of R_1 about one-tenth or so of R_2. Knowing the equilibrium value of the voltage drop across R_1, one knows the value of I_{CQ2}/I_{CQ1} in terms of *p*, from equation 5. If V_1 is 60mV, and *p* unity, I_{CQ2} will be one-tenth I_{CQ1}. Assuming the potential drop across R_3 to be substantially all attributable to I_{CQ2} and to be substantially equal to V_2, one can calculate R_3 by Ohm's Law to be V_2/I_{CQ2}, which equals $(V_2/I_{CQ1})(I_{CQ1}/I_{CQ2})$, which equals $R_2(I_{CQ1}/I_{CQ2})$ or about **10** R_2. Such calculations yield values of R_1, R_2 and R_3 of 600, 5600 and 56000 ohms, respectively, for example, with R_4 chosen to supply an I_{CQ1} of 0.1mA, an I_{CQ2} of 0.01mA, and an I_{CQ3} of 0.1mA—i.e., a total of some 0.2mA.

The FIG. **1** reference potential generator has the shortcoming, acceptable in some applications but not in others, that it depends upon V_{BEQ3} being determinate to obtain good regulation of V_{21}. V_{BEQ3} changes by 18 millivolts for each doubling of its collector current, however, so if the current applied between T_1 and T_2 of the reference voltage generator changes, the regulation of V_{21} will be affected. An improvement would be to provide a threshold voltage for sensing the potential between T_1 and the second end of R_3 that would be substantially less dependent upon the operating current supplied to the reference potential. It would also be desirable, if possible, to reduce the current loading upon T_3 posed by the shunt regulating device while at the same time increasing the transconductance of the shunt regulating device.

The present inventor observed that the regulated value of V_{21} applied to the series combination of R_2 and self-biased Q_1 causes the collector current I_{CQ1} of transistor Q_1 to be quite well-regulated so the value of V_{BEQ2} is substantially independent of the operating current supplied to the reference potential generator of FIG. **1**. FIG. **2** shows a reference potential generator taking advantage of this observation to provide improvements upon the FIG. **1** reference potential generator.

In FIG. **2**, a differential input amplifier A_1, such as an operational amplifier, replaces Q_3 in combination with R_4 to provide the means for sensing when the potential between T_1 and T_3 exceeds a predetermined threshold value to generate a reference potential directly related to such excess. The threshold value is set by V_{BEQ1}, which because of V_{21} being regulated is of more determinate value than V_{BEQ3}. Rather than measuring the potential between T_1 and T_3 directly, one does it indirectly by comparing the potentials between the base of Q_1 and T_3. This permits substantially greater freedom of design of the amplifier T_3 works into. A_1 may use Darlington transistors of FET's in its input stage to reduce loading on the base of Q_1 and on T_3, and one may readily use cascaded amplifier stages to secure very high transconductance in A_1 to improve the regulation of V_{41}.

FIG. 3 shows a reference potential generator that may be used instead of the FIG. 2 reference potential generator, in which V_{BEQ2} rather than V_{BEQ1} is used as the threshold value against which the potential at T_3 is compared. R_3' is equal to $R_3(R_1 + R_2)/R_1$. Other modifications are possible in which the threshold value is between V_{BEQ1} and V_{BEQ2}, being obtained from a point along R_1. Modifications of the FIG. 2 reference potential generator in which the inputs of A_1 are taken from taps on resistors R_2 and R_3 are also possible.

FIG. 4 shows a modification that can be made to any of the reference potential generators shown in FIGS. 1 through 3, which modification will increase the reference potential V_{41} it produces by a factor m. This modification consists of a potential divider D_1 having an input terminal connected to T_4 and an output terminal connected to T_2. Potential divider D_1 divides the potential V_{41} by a factor m to obtain the potential V_{21} for application between T_1 and T_2.

FIGS. 5 and 6 show modifications of the reference potential generators of FIGS. 2 and 3, respectively, useful for providing V_{24} reference potentials relatively negative, rather than relatively positive, as referred to a fixed potential shown as ground.

FIG. 7 shows a modification that can be made to either of the reference potential generators shown in FIGS. 5 and 6, which modification will increase the reference potential V_{24} it produces by a factor m. This modification consists of a potential divider D_2 having an input terminal connected to T_4 and an output terminal connected to T_1. Potential divider D_2 divides the potential V_{24} by a factor m to obtain the potential V_{21} for application between T_1 and T_2.

In the circuits of FIGS. 2, 3, 5 and 6 as shown or as modified by FIGS. 4 and 7, R_4 may be omitted if A_1 is a conventional operational amplifier rather than an operational transconductance amplifier.

In the reference potential generators of the sort shown in FIGS. 2, 3, 5 and 6, the value of V_{21} that exhibits a zero temperature coefficient will depart somewhat from $V_{g(0)}$ depending upon the temperature coefficient of the resistors R_1, R_2 and R_3. The $(V_{21} - V_{BEQ1})$ drop across R_2 of about 600mv will increase 1.75mV per Kelvin increase in temperature due to the negative temperature coefficient of V_{BEQ1}. So I_{CQ1}, the major portion of the current through R_2, will be held substantially constant if R_2 has a positive temperature coefficient as expressed in percentage equal to that of the potential drop across it $+1.75mV/k/600mv = +0.29\%/K$. Such temperature coefficients can be achieved with ion-implanted integrated resistors. But diffused resistors normally have lower positive temperature coefficients—e.g., $+0.2\%/K$—causing the zero-temperature-coefficient value of V_{21} to vw less than $V_{g(0)}$ by thirty-five millivolts or so.

While the provision of a zero-temperature-coefficient reference potential V_{41} (or V_{24}) equal to $V_{g(0)}$ has been specifically treated in the foregoing specification, the reference potential generator configurations shown are useful for generating reference potentials having other temperature coefficients. These V_{41}'s (or V_{24}'s) may be negative-temperature-coefficient potentials that are a multiple of V_{21}'s that range between V_{BEQ1} to $V_{g(0)}$. Or these V_{41}'s (or V_{24}'s) may be positive-temperature-coefficient potentials that are multiples of V_{21}'s larger than $V_{g(0)}$.

What is claimed is:

1. A reference potential generator comprising:
 first and second and third terminals;
 bias means for tending to increase the potential between said first and said second terminals;
 first and second transistors of the same conductivity type, each having base and emitter electrodes with a base-emitter junction therebetween and having a collector electrode, each of their emitter electrodes being directly connected without substantial intervening impedance to said first terminal;
 a first resistive element having a first end which connects to the base electrode of said first transistor and having a second end which connects to the base electrode of said second transistor and has the collector electrode of said first transistor connected thereto;
 a second resistive element having a first end connected to said second terminal and having a second end connected to the first end of said first resistive element;
 a third resistive element having a first end connected to said second terminal and having a second end connected to a third terminal and to the collector electrode of said second transistor;
 means for sensing when the potential between said first and third terminals exceeds a predetermined threshold value to decrease the potential between said first and said second terminals, thereby to generate a reference potential; and
 means applying between said first and said second terminals a fixed portion of said reference potential, thereby completing a feedback loop for regulating said reference potential to prescribed value.

2. A reference potential generator as set forth in claim 1 wherein said means for sensing when the potential between said first and said third terminals exceeds a predetermined threshold potential to generate a reference potential directly related to said excess senses the potential between said first and said third terminals directly and comprises:
 a third transistor of said same conductivity type having emitter and base electrodes respectively connected to said first terminal and to said third terminal, having a base emitter junction between its emitter and base electrodes, the offset potential of which corresponds to said predetermined threshold value, and having a collector electrode direct coupled to said second terminal.

3. A reference potential generator as set forth in claim 1 wherein said means for sensing when the potential between said first and said third terminals exceeds a predetermined threshold potential to generate a reference potential directly related to said excess senses the potential between said first and said third terminals indirectly and comprises:
 a differential-input amplifier having an inverting input terminal connected to said third terminal, having a non-inverting input terminal to which a predetermined threshold potential related to at least one of the base potentials of said first and said second transistors is applied, and having output terminals between which said reference potential is supplied.

4. A solid-state temperature-compensated voltage supply comprising:

first and second transistors;

a resistor connected between the base of said first transistor and the base of said second transistor;

circuit mears for furnishing supply voltage to said two transistors to develop current flow therethrough with a current through said first transistor also flowing through said resistor;

means for sensing the magnitude of the respective currents flowing through said two transistors;

voltage-control means responsive to the currents sensed by said sensing means and operable to adjust the emitter potentials of said transistors to maintain the magnitude of said transistor currents at levels which provide a predetermined non-unity ratio of current densities within the two transistors responsive to which they exhibit a difference in their emitter-to-base offset potentials that is applied to said resistor connected between their bases to cause the current through said resistor to vary positively with respect to the temperature of said two transistors;

means for developing a first voltage proportional to said resistor current and for combining said first voltage with a second voltage which varies negatively with respect to said temperature to produce a combined voltage having minimal overall variation with respect to said temperature; and

output means coupled to said last named means and including an output terminal providing an output voltage proportional to said combined voltage.

5. A voltage supply as claimed in claim **4**, wherein said voltage-control means comprises:

a high-gain amplifier serving as a comparator responsive to signals proportional to said current flows through said first and said second transistors to produce an output signal corresponding to the difference between said signals proportional to said current flows; and

means coupling a voltage proportional to said output signal to the emitters of said transistors to drive the emitter potentials to values providing the desired ratio of current density in said transistors.

6. A voltage supply as claimed in claim **5** wherein said sensing means comprises first and second load resistors connected in the collector circuits of said first and said second transistors, respectively.

7. A voltage supply as claimed in claim **4** wherein the emitters of said first and said second transistors are connected together to provide equal emitter potentials.

* * * * *